An Iterative Method for ASC Hybrid Precoding Structure for mmWave Ma-MIMO Systems

Amirreza Moradi[1], K. Mohamed-pour[2], Nasim Jafari Farsani[3]
1- Department of Electrical Engineering, K. N. Toosi University of Technology Tehran, Iran.
Email: Amirreza_moradi@email.kntu.ac.ir
2- Department of Electrical Engineering, K. N. Toosi University of Technology Tehran, Iran.
Email: kmpour@kntu.ac.ir
3- Department of Electrical Engineering, K. N. Toosi University of Technology Tehran, Iran.
Email: jafari_nasim@email.kntu.ac.ir

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ABSTRACT:
The use of millimeter wave (mmWave) massive multiple input multiple output (Ma-MIMO) systems makes it possible to meet the essential needs of future generation wireless systems and solve the impending wireless network crisis. The mmWave Ma-MIMO Technique offers higher numbers of antennas and carrier frequencies. Hybrid precoding is considered as a key technique for the practical deployment of mmWave Ma-MIMO systems, since it significantly decreases the implementation costs, energy consumption, and hardware complexity. The large using of mmWave Ma-MIMO technologies in future generation wireless systems, causes imperative develop cost-effective hybrid precoding solutions that match the various application cases of these systems. The fully-connected structure can offer spectral efficiency (SE) close to the fully-digital precoding but, unfortunately with high energy consumption. Furthermore, the sub-connected structure with reduced power consumption, provides poor SE. Therefore, the trade-off between SE and energy efficiency (EE), can be made, and in this paper, we consider an adaptive sub-connected (ASC) hybrid precoding structure, where a switch network is able to provide dynamic connections from phase shifters to radio frequency (RF)chains. The simulation results indicate that in terms of SE, the proposed algorithm with ASC structure obtains higher performance than the sub-connected structure. As a result, since the ASC structure reduces the number of phase shifters, it can offer a better EE compared to the sub-connected structure.

KEYWORDS: ASC Structure; Energy Efficiency; Hybrid Precoding; mmWave Ma-MIMO; Spectral Efficiency.

1. INTRODUCTION

Increasing the number of antennas at the base station has made traditional linear precoding schemes such as Zero Forcing and Minimum Mean Square Error able to attain near-optimal performance obtained from dirty paper coding in downlink communication [1],[2],[3]. However, in traditional designs, the more antennas are used, the more RF chains will be required. A large number of RF chains in Ma-MIMO systems leads to huge costs and energy consumption[4]. To handle the mismatch between the number of RF chains and antennas, hybrid precoding schemes including, cost-effective variable phase shifters, are introduced. Hybrid precoding schemes employ a phase-only RF and a baseband precoder in the analog and in the digital domain, respectively[5],[6]. Although hybrid precoding architecture reduce**s** energy consumption and increase**s** EE, it degrades performance. Efficient hybrid precoding schemes should be developed to acquire the best trade-off between performance and hardware complexity.

Moreover, the millimeter wave (mmWave) frequencies have been put forward as prime candidates for future generation cellular systems[7],[8]. Fortunately, the reduction in wavelength of mmWave MIMO systems allows the use of large-scale antenna arrays in transmitters, which can provide significant benefits in beamforming to overcome path loss[9]. The hybrid baseband and RF processing are particularly appropriate for mmWave MIMO systems because the mmWave systems rely heavily on RF processing, and the hybrid processing can effectively reduce the excessive cost of RF chains[10]. In order to further reduce the hardware complexity, the number of phase shifters in use can also be reduced. Therefore, based on how each RF chain is connected to the antennas, the

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hybrid precoding architectures can be classified into the fully-connected and sub-connected architectures[11].

In the fully-connected structure, every separate RF chain is connected to all antennas, which can achieve excellent precoding performance[12]. The downside of this structure is the number of the required phase shifters, which is equal to the product of the number of RF chains and antennas. Since the number of phase shifters is still high, the problem with high cost, complexity, and power consumption remains[13].

In the sub-connected structure, each RF chain is linked to only a subset of antennas, which reduces the number of phase shifters and hardware cost, complexity, and power consumption. Although the sub-connected architecture sacrifices some beamforming gain. Based on the successive interference cancellation (SIC),the authors in [14] proposed an iterative hybrid precoding algorithm. Also, based on alternating minimization between digital and analog precoders, the authors in [13] have proposed SDR-ALTMIN algorithm. Which it offers optimal solutions for both subproblems of analog and digital precoders in each alternating iteration.

To reduce the power consumption and hardware complexity of phase shifters, in recent years, on-off binary state switches have been proposed for hybrid precoding. However, the poor SE offered by switch-based hybrid precoding structures is disappointing.

To decrease the number of phase shifters and complexity, dynamic hybrid precoding is proposed. This paper considers an (ASC) hybrid precoding structure[15], where the antenna elements are dynamically divided into many subsets, and by the switch network, every subset is connected to all the RF chains with a few numbers of phase shifters.

In summary, the major contributions of this paper are as follows:

- ASC structure for hybrid precoding is proposed, which, by using an analog switch network, each RF chain is able to be dynamically connected to a subset of antennas via phase shifters.
- Switch, and digital precoding matrix is derived by using an alternative optimization. Moreover, we calculate the phase shifter matrix through an iterative solution. Then to satisfy the unit module constant we use the element-wise normalization.
- In the end, we compare the various modes of the proposed hybrid precoding algorithm with the SDR-Altmin algorithm and SIC-based precoder in terms of SE and EE. Based on the simulation results, the proposed hybrid precoding based on the ASC structure can achieve proper SE by a huge reduction in power consumption.

The remaining of this manuscript is organized as follows. The hybrid precoding system model and the mmWave channel assumption are introduced in Section 2; In section 3, the hybrid precoding formulation of the proposed structure is described, and next, the hybrid precoding algorithm and the EE performance are elaborately discussed in section 4. Simulation results are evaluated in Section 5, and finally, in Section 6 the conclusion is summarized.

The following notation is used throughout this paper: $\mathbb{C}$ is the set of complex numbers; $\Re\{\cdot\}$ shows the real part of complex variable; $\boldsymbol{X}$ is a matrix; $\boldsymbol{x}$ is a vector. $\boldsymbol{X}^{\mathrm{H}}$ denotes the conjugate transpose of $\boldsymbol{X}$; $\boldsymbol{X}_{i,j}$ is the entry in the ith row and jth column of $\boldsymbol{X}$; $\|\boldsymbol{X}\|_F$ shows the Frobenius norm of $\boldsymbol{X}$; $(\cdot)$denotes the trace/Frobenius product ; $\boldsymbol{X}^{-1}$ and $\boldsymbol{X}^{\dagger}$express the inverse and the Moore-Penrose pseudo inverse of matrix $\boldsymbol{X}$; Expectation is denoted by $\mathbb{E}[\cdot]$; $\mathrm{tr}(\boldsymbol{X})$ is the trace of matrix $\boldsymbol{X}$; $\oslash$ stands for element-wise division; $\mathbb{1}(\cdot)$ represents the indicator function; $\mathbf{1}_{M\times N}$ denotes an $M\times N$ matrix of ones. $\mathrm{XN}(x,y)$ denotes the complex Gaussian distribution with x and y as the mean and covariance, respectively; $\boldsymbol{I}_N$ is the $N\times N$ identity matrix.

2. SYSTEM MODEL

The ASC hybrid precoder design with dynamic structure for Single-user mmWave Ma-MIMO system demonstrates in Fig.1. The base station is equipped with N_t transmitting antennas, also on the receiver side, we suppose a single user has N_r receiving antennas. Antennas are shaped as a uniform plane array (UPA) to transmit N_s data streams. The transmitter and receiver are equipped with multiple RF chains to ensure transferring multiple data streams in a parallel mode. The number of RF chains are shown as N_{RF}, and it satisfies $N_s \le N_{RF} \le min\{N_t.N_r\}$ constraints.

Mathematically, the transmitted signal $N_t \times 1$ vector is written as $\boldsymbol{x} = \boldsymbol{F}_S \boldsymbol{F}_{PS}\boldsymbol{F}_{BB}\boldsymbol{s}$. Also, $\mathbf{s} = [s_1,\ldots,s_{N_s}]^{\mathrm{T}}$ denotes the $N_s\times 1$ transmitted signal vector, and the average transmitted power of the Gaussian data symbol can be expressed as $\mathbb{E}\left[\boldsymbol{s}\boldsymbol{s}^{\mathrm{H}} = \frac{\boldsymbol{I}_{N_S}}{N_S}\right]$.

The hybrid precoding structure contains of three processing stages: digital precoder $\boldsymbol{F}_{BB} \in \mathbb{C}^{N_{RF}\times N_S}$, followed by an analog phase shifter precoding matrix $\boldsymbol{F}_{PS} \in \mathbb{C}^{N_A\times N_{RF}}$ and an analog switch precoding matrix$\boldsymbol{F}_S \in \mathbb{C}^{N_t\times N_A}$. Also N_A denotes the number of adders. The following expression indicates the received signal after decoding processes

$$y = \sqrt{\rho}\,(\boldsymbol{W}_{RF}\boldsymbol{W}_{BB})^{\mathrm{H}}\boldsymbol{H}\boldsymbol{F}_S\,\boldsymbol{F}_{PS}\boldsymbol{F}_{BB}\boldsymbol{s} + (\boldsymbol{W}_{RF}\boldsymbol{W}_{BB})^{\mathrm{H}}\,\mathbf{n} \qquad (1)$$

where ρ and $\boldsymbol{H}$ denote the average received power and $N_r \times N_t$ scattering channel matrix, respectively. On the receiver side decoding process is similar to the precoding at the transmitter. Therefore, we have a low dimension

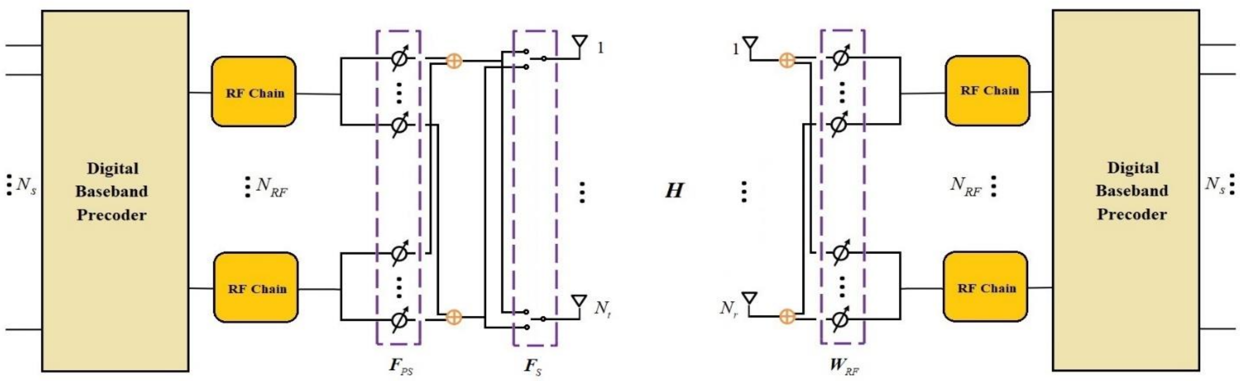

Fig. 1. Hybrid precoder design with dynamic structure

digital (baseband) combiner $\boldsymbol{W}_{BB} \in \mathbb{C}^{N_{RF} \times N_S}$,and an analog RF combining matrix $\boldsymbol{W}_{RF} \in \mathbb{C}^{N_r \times N_{RF}}$. Noise vector $\mathbf{n}$, which has dimension of $N_r \times 1$ is assumed to be independent and identically distributed random variable (i.i.d) with zero mean and average power σ_n^2, i.e., $\mathbf{n} \sim \mathrm{XN}(0\,,\,\sigma_n^2)$.

According to (1), the SE obtained by hybrid precoding system can be expressed as

$$\mathrm{R} = \log_2 \left| \boldsymbol{I}_{N_S} + SINR \right| \tag{2}$$

where SINR denotes the signal-to-noise-plus-interference ratio and recasts as in

$$\mathrm{SINR} = \frac{\rho}{\sigma_n^2 N_S} \cdot \left| \left(\boldsymbol{W}^{\mathrm{H}}\boldsymbol{W}\right)^{-1} \boldsymbol{W}^{\mathrm{H}}\boldsymbol{H}\boldsymbol{F}(\boldsymbol{F})^{\mathrm{H}}\boldsymbol{H}^{\mathrm{H}}\boldsymbol{W} \right| \tag{3}$$

Where $\boldsymbol{F}=\boldsymbol{F}_S\,\boldsymbol{F}_{PS}\boldsymbol{F}_{BB}$ and $\boldsymbol{W}=\boldsymbol{W}_{RF}\boldsymbol{W}_{BB}$.

Then, we substitute both of (3) and definition noise covariance matrix $\boldsymbol{R}_n = \sigma_n^2 \boldsymbol{W}^{\mathrm{H}}\boldsymbol{W}$ into (2). Therefore, the SE expression with regard to precoding matrices can be rewrite as in (4).

$$\mathrm{R} = \log_2 \left| \boldsymbol{I}_{N_S} + \frac{\rho}{N_S} \boldsymbol{R}_n^{-1} \boldsymbol{W}^{\mathrm{H}}\boldsymbol{H}\boldsymbol{F}\boldsymbol{F}^{\mathrm{H}}\boldsymbol{H}^{\mathrm{H}}\boldsymbol{W} \right| \tag{4}$$

In this paper, we assume that the channel state information (CSI) is perfectly-known for both transceiver sides. Designing the unconstrained optimum digital precoders (combiners), $\boldsymbol{F}_{opt}$ $(\boldsymbol{W}_{opt})$, is usually define by using the channel singular value decomposition (SVD) [16][17].

$$\begin{aligned} \boldsymbol{F}_{opt} &= \boldsymbol{V}_{:,1:N_S} \\ \boldsymbol{W}_{opt} &= \boldsymbol{U}_{:,1:N_S} \end{aligned} \tag{5}$$

Where $\boldsymbol{U}$ and $\boldsymbol{V}^{\mathrm{H}}$are left and right singular matrices.

The traditional channel model is not suited for mmWave Ma-MIMO systems due to the characteristics of extreme pathloss and high level of correlation between antennas. As the result, the narrowband clustered channel modeling based on the Saleh-Valenzuela is adopted [18] to describe the propagation channel between the transmitter and receiver. In this model, we assume the channel matrix $\boldsymbol{H}$ contains of N_{cl} clusters, and each cluster combined of N_{ray} rays. Consequently, the total number of propagation paths would be $L = N_{cl}N_{ray}$, therefore the channel matrix $\boldsymbol{H}$ can be can be written as

$$\mathbf{H} = \sqrt{\frac{N_t N_r}{L}} \sum_{i=1}^{N_{cl}} \sum_{k=1}^{N_{ray}} \alpha_{ik} \mathbf{a}_r\left(\phi_{ik}^r, \theta_{ik}^r\right) \mathbf{a}_t^H\left(\phi_{ik}^t, \theta_{ik}^t\right) \tag{6}$$

Where α_{ik} denotes the complex gain of the kth ray in the ith scattering cluster. the azimuth angle of departure (AOD) and angle of arrival (AOA) and the elevation AOD and AOA associated with the kth ray in ith cluster denotes by $\emptyset_{ik}^t, \emptyset_{ik}^r, \theta_{ik}^t$ and θ_{ik}^r, respectively. Moreover, $\mathbf{a}_t(\emptyset_{ik}^t, \theta_{ik}^t)$and $\mathbf{a}_r(\emptyset_{ik}^r, \theta_{ik}^r)$ indicates the vector of the array response of the transmitter and receiver, respectively[19].

The array response vector for an UPA with dimension of $M \times N$ recast as

$$\boldsymbol{a}_{\mathrm{UPA}}(\emptyset, \theta) = \frac{1}{\sqrt{MN}}\left[1 \ \ldots \ e^{j\pi(m\sin\emptyset\sin\theta + n\cos\theta)} \ \ldots \ e^{j\pi((M-1)(\sin\emptyset\sin\theta + (N-1)\cos\theta)}\right]^T$$

Where $0 \le m \le M-1$ and $0 \le n \le N-1$ and inter-element space is assumed to be half-wavelength.

3. PROBLEM FORMULATION

Based on the ASC structure, the corresponding hybrid precoding problem can be approximately formulated as [4]

$$\begin{aligned}&\min_{F_S F_{PS} F_{BB}} \left\| F_{opt} - F_S F_{PS} F_{BB} \right\|_F^2 \\ &\quad \text{s.t.} \left[F_S \right]_{mn} \in \{0,1\} \quad \forall m,n \\ &\qquad \left\| F_{S,m,:} \right\|_0 = 1 \quad \forall m \\ &\qquad \left| F_{PS,ij} \right| = 1 \quad \forall i,j \\ &\qquad \left\| F_S F_{PS} F_{BB} \right\|_F^2 = N_s \end{aligned} \tag{7}$$

Also, the mathematical formulation of the combining problems can be expressed as follow.

$$\begin{aligned}&\min_{W_{RF} W_{BB}} \left\| W_{opt} - W_{RF} W_{BB} \right\|_F^2 \\ &\quad \text{s.t.} \left| W_{RF,bd} \right| = 1 \quad \forall b,d \end{aligned}$$

(8)

To assure the unit modulus constraints, each entry of F_{PS} and W_{RF} have constraint of

$$\begin{aligned}&\left| F_{PS,ij} \right| = 1 \quad \forall i,j \\ &\left| W_{RF,bd} \right| = 1 \quad \forall b,d \end{aligned} \tag{9}$$

Furthermore, Switch precoding matrix have binary constraint $[F_S]_{mn} \in \{0,1\}$ Finally, $\|F_S F_{PS} F_{BB}\|_F^2 = N_S$ determines the total transmit power constraint.

The following of this paper will only consider the hybrid precoding problem, and for the sake of simplicity and to avoid additional calculations, the problem on the receiver can be solved in a similar way as [20] and[21]. Sadly, the binary constraint of the analog switch network F_S and the equal amplitude of the phase matrix F_{PS} will make the problem (7) non-covex, therefore, it is extremely difficult to jointly optimize the hybrid precoder over F_S, $F_{PS}$and F_{BB} to obtain the optimal solution of (7). Therefore, In the continuation of this paper, a hybrid precoding algorithm for further analysis of the optimization problem will be described.

4. HYBRID PRECODING STRATEGY

4.1. SWITCH PRECODING MATRIX OPTIMIZATION

By assuming that F_S and F_{BB} are fixed, the optimization problem of F_S can be expressed as

$$\begin{aligned}&\min_{F_S} \left\| F_{opt} - F_S F_{PS} F_{BB} \right\|_F^2 \\ &\quad \text{s.t.} \left[F_S \right]_{mn} \in \{0,1\} \quad \forall m,n \\ &\qquad \left\| F_{S,m,:} \right\|_0 = 1 \quad \forall m \end{aligned} \tag{10}$$

Let $F_{eff} = F_{PS} F_{BB}$. The problem (5) can be rewritten as

$$\begin{aligned}&\min_{F_S} \left\| F_{opt} - F_S F_{eff} \right\|_F^2 \\ &\quad \text{s.t.} \left[F_S \right]_{mn} \in \{0,1\} \quad \forall m,n \\ &\qquad \left\| F_{S,m,:} \right\|_0 = 1 \quad \forall m \end{aligned} \tag{11}$$

We know $\|F_{opt}^H\text{-}F_{eff}^H F_S^H\|_F^2 = \sum_{m=1}^{N_t} \left\| f_{opt,m}^H F_{eff}^H f_{S,m}^H \right\|_2^2$ Therefore, the problem (11) can be divided to N_t independent subproblems

$$\begin{aligned}&\min_{fm} \left\| f_{opt,m}^H - F_{eff}^H f_{S,m}^H \right\|^2 \\ &\quad \text{s.t. } f_{S,m}(n) \in \{0,1\} \quad \forall m,n \\ &\qquad \left\| f_{S,m} \right\|_0 = 1 \quad \forall n \end{aligned} \tag{12}$$

where $f_{opt,m}^H \in \mathbb{C}^{1\times N_s}$ and $f_{S,n}^H \in \mathbb{C}^{1\times N_A}$ are the mth row of F_{opt} and F_S, respectively and $f_{S,m}(n)$ denotes the nth entry in $f_{S,m}$.

Due to the characteristic of F_S, problem (12) can be considered as a sparse approximation problem. It's noteworthy that vector $f_{S,m}$ has only an one non-zero entry. Therefore, by using exhaustive search; it is possible to provide the sparse approximation of the measurement vector $f_{opt,m}^H$ under the baseband precoding dictionary f_{eff}^H [15],[22]. Thus, the problem in (10) can have a solution as follow

$$\left[F_S \right]_{mn} = \begin{cases} 0, & n \neq n_m^{opt}, \\ 1, & n = n_m^{opt}, \end{cases} \tag{13}$$

Where $n_m^{opt} = \arg\min \left\| f_{opt,m}^H - f_{eff,n_m}^H \right\|_2^2$ and f_{eff,n_m}^H indicates the n_mth row of F_{eff}.

4.2. ANALOG PHASE SHIFTER PRECODING MATRIX OPTIMIZATION

In this subsection, by fixing the optimized analog switch precoding matrix F_S and the digital precoding matrix F_{BB}, we optimize the analog phase shifter precoding matrix F_{PS}.

$$\begin{aligned}&\min_{F_{PS}} \left\| F_{opt} - F_S F_{PS} F_{BB} \right\|_F^2 \\ &\quad \text{s.t.} \left| F_{PS,ij} \right| = 1 \quad \forall i,j \end{aligned} \tag{14}$$

Proposition 1. the optimal solution of (15) is given by

$$F_{PS} = \left(F_S^H F_{opt} F_{BB}^H - F_S^H F_S F_{PS} F_{BB} F_{BB}^H \right) + F_{PS} \tag{15}$$

Proof: The proof is provided in Appendix A. ■

In the proposed method, we considered F_{PS} is as a unit modulus matrix, i.e., $|F_{PS,mn}| = 1$; therefore, an element-wise normalization is proposed to satisfy unit modulus constraints.

$$F_{PS} = F_{PS} \oslash (|F_{PS}|) \tag{16}$$

4.3. DIGITAL PRECODING MATRIX

OPTIMIZATION

The digital precoding matrix $\boldsymbol{F}_{BB}$ optimization problem can be stated as

$$\min_{\boldsymbol{F}_{BB}} \left\| \boldsymbol{F}_{opt} - \boldsymbol{F}_S \boldsymbol{F}_{PS} \boldsymbol{F}_{BB} \right\|_F^2 \tag{17}$$

A suboptimal solution to the problem (17) can be obtained by least squares

$$\boldsymbol{F}_{BB} = \left(\boldsymbol{F}_S \boldsymbol{F}_{PS} \right)^{\dagger} \boldsymbol{F}_{opt} \tag{18}$$

Finally, to satisfy the constraint on the transmit power $\|\boldsymbol{F}_{PS}\boldsymbol{F}_S\boldsymbol{F}_{BB}\|_F^2 = N_s$, we normalize the digital precoder $\boldsymbol{F}_{BB}$ i.e.

Algorithm 1 Proposed Hybrid precoding

Input: $\boldsymbol{F}_{opt}$

Initialization: $\boldsymbol{F}_{PS}^{(0)}$ **and** $\boldsymbol{F}_{BB}^{(0)}$ **are generated randomly,** $t = 0$

Repeat

1. $t = t + 1$**;**
2. **Fix** $\boldsymbol{F}_{BB}^{(t-1)}$ **and** $\boldsymbol{F}_{PS}^{(t-1)}$**, Optimize** $\boldsymbol{F}_S^{(t)}$ **according to (13);**
3. **Update** $\boldsymbol{F}_{BB}^{(t)}$ **with (18) by fixing** $\boldsymbol{F}_{PS}^{(t-1)}$ **and** $\boldsymbol{F}_S^{(t)}$**;**
4. **Fix** $\boldsymbol{F}_{BB}^{(t)}$ **and** $\boldsymbol{F}_S^{(t)}$ **Optimize** $\boldsymbol{F}_{PS}^{(t)}$ **by (15);**
5. **Element-Wise Normalization:** $\boldsymbol{F}_{PS}^{(t)} = \boldsymbol{F}_{PS}^{(t)} \oslash \left(\left| \boldsymbol{F}_{PS}^{(t)} \right| \right)$

Until convergence

6. $\boldsymbol{F}_{BB} = \frac{\sqrt{N_s}}{\|\boldsymbol{F}_{PS}\boldsymbol{F}_S\boldsymbol{F}_{BB}\|_F} \boldsymbol{F}_{BB}$

Output: $\boldsymbol{F}_{BB}$, $\boldsymbol{F}_S$, $\boldsymbol{F}_{PS}$

$$\boldsymbol{F}_{BB} = \frac{\sqrt{N_s}}{\|\boldsymbol{F}_{PS}\boldsymbol{F}_S\boldsymbol{F}_{BB}\|_F} \boldsymbol{F}_{BB} \tag{19}$$

Table 1. states the pseudo-code for the proposed hybrid precoder $\boldsymbol{F}=\boldsymbol{F}_S \boldsymbol{F}_{PS}\boldsymbol{F}_{BB}$ solution.

4.4. ENERGY EFFICIENCY

In this subsection, the EEs of the ASC hybrid precoding structure and other relevant works are formulated. According to [14], [23], the EE of a communications system is specified by the total power consumption and the SE. The EE in mmWave MIMO systems is represented by η, and can be written as

$$\eta = \frac{R}{P_{total}} \tag{20}$$

Where R indicates the hybrid precoding achievable SE, and P_{total} denotes the power consumption of each algorithm which can be represented as

$$P_{total} = \begin{cases} P_t + P_{BB} + P_{RF} N_{RF} + P_{PS} N_{RF} N_A + P_S N_t, & \text{PRO} \\ P_t + P_{BB} + P_{RF} N_{RF} + P_{PS} N_t, & \text{SDR-Altmin} \\ P_t + P_{BB} + P_{RF} N_{RF} + 2P_{PS} N_t, & \text{SIC} \end{cases} \tag{21}$$

Where P_t shows the transmit power, P_{BB}, P_{RF} , P_{PS} , and P_S indicate the power consumed by the baseband, each RF chain, phase shifter, and switch, respectively.

5. SIMULATION RESULTS

In this section, we assess the SE and the EE performance of the proposed hybrid precoding algorithm by numerical simulations, where the SDR-Altmin and SIC algorithms are taken into account as the competitors. The environment for mmWave propagation is assumed to contains of $N_{cl} = 5$ clusters, and each cluster combined of $N_{ray} = 10$, i.e., the total path number equals to $L = 50$. Both the transmitter and receiver are equipped with UPA. For each cluster, the average azimuth and elevation AOD (AOA) is drawn independently from uniform distribution over $[0,2\pi)$; In clusters, a 10-degree angular spread Laplace distribution conforms the azimuth and elevation AOD (AOA). Each path is assumed to have complex gain of $\mathbf{n} \sim \mathcal{CN}(0,1)$. For evaluating the EE simulation, parameters are set as follows:
$P_t = 1\text{W}$, $P_{BB} = 0.2\text{W}$, $P_{PS} = 0.04\text{W}$, $P_{RF} = 0.250\text{W}$, and $P_S = 0.005\text{W}$ [14], [24].

All of the reported results are achieved by calculating the average over 1000 independent channel realizations.

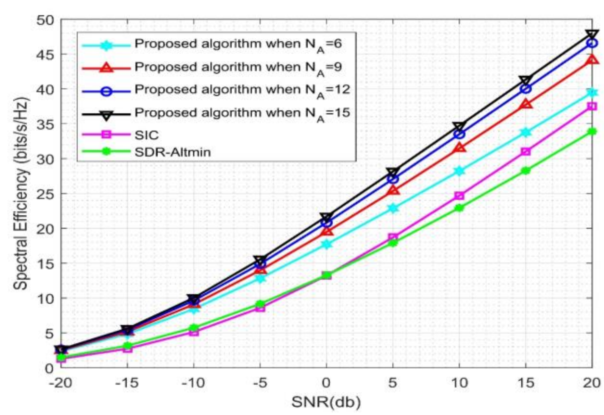

Fig. 2. Spectral efficiencies achieved against SNR

Fig. 2. shows the SEs of all schemes versus SNR where $N_t = 12 \times 12$, $N_r = 4 \times 4$, $N_s = 4$, $N_{RF} = 4$. It can be seen the SE of the proposed algorithm increases as the number of adders grows. At high SNRs, the performance of the proposed algorithm when $N_A = 6$ is not favorable compared to other cases of the proposed

algorithm. Also, the proposed algorithm offers higher SE in all cases than SIC and SDR-Altmin algorithm. Nevertheless, the SIC algorithm reduces its distance from the proposed novel algorithm when $N_A = 6$.

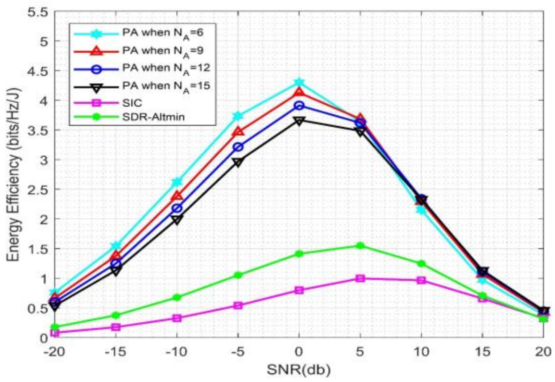

Fig. 3. Energy efficiencies achieved against SNR

Fig. 3 plots the EEs against the SNR where $N_t = 12 \times 12$, $N_r = 4 \times 4$, $N_s = 4$, $N_{RF} = 4$. Evaluating meticulously, the proposed algorithm shows a huge improvement compared to other algorithms. The EE of our algorithm grows significantly as the SNR increases until when $\mathrm{SNR} = 0\mathrm{dB}$ in which the proposed algorithm has its best performance. Since an increase in the number of adders means excess in the number of phase shifters and switches, it leads to an increase in power consumption, which will lead to a decrease in EE. The performance of other methods decreases when $\mathrm{SNR} > 5\mathrm{dB}$.

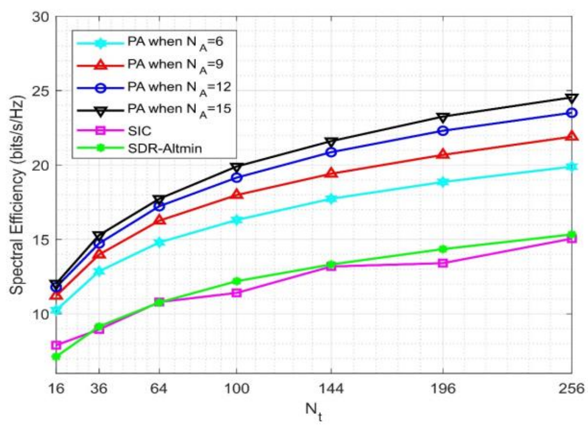

Fig. 4. Spectral efficiencies achieved against N_t

Fig.4 and Fig 5 draw the comparisons between the SE and the EE performance vs. the number of the transmitter antennas N_t when $N_r = 4 \times 4$, $N_s = 4$, $N_{RF} = 4$, $\mathrm{SNR} = 0\mathrm{dB}$. The performance of the SIC and SDR-Altmin algorithm is almost the same except when $N_t = 100$ and $N_t = 196$, which SDR-Altmin performs slightly better. The SE difference between the proposed algorithm when $N_A = 6$ and those SIC and SDR-Altmin algorithms is about *4Bits/s/Hz* to *5Bits/s/Hz.* Also, the SE of the proposed algorithm enhances as the number of adder increase. In case when $N_A = 15$, the proposed algorithm has about *4Bits/s/Hz* to *5Bits/s/Hz* SE gain for when $N_t > 64$

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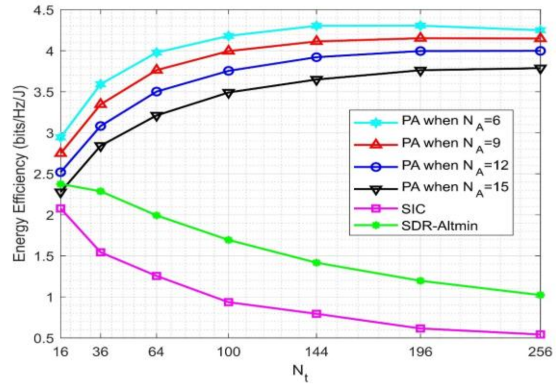

Fig. 5. Energy efficiencies achieved against N_t

In Fig.5 the EEs of competitor methods decrease with N_t. It can be seen that the SIC algorithm shows a very poor EE, and SDR-Algorithm performs about *0.5Bits/Hz/J* better compare to SIC-algorithm. The EE of the proposed algorithm in all cases increases with N_t until when $N_t = 196$. Also, the EE of the different proposed algorithm modes decreases as the number of adders grows.

6. CONCLUSION

This paper concentrates on designing a low-complexity hybrid precoder algorithm for ASC structure. A hybrid precoding algorithm for a single-user mmWave Ma-MIMO communication systems is introduced. The baseband, analog switch, and phase shifter precoding matrices are optimized via an alternative and iterative solution. Simulation results have indicated that in terms of SE, the proposed algorithm can offer much higher SE in comparison to those of the sub-connected structure SIC and SDR-Altmin algorithm with significantly fewer phase shifters. Moreover. Considering the low number of phase shifters in the proposed algorithm, a remarkable EE improvement is provided by the proposed algorithm.

7. APPENDIX A
PROOF OF PROPOSITION 1

The objective function in (8) can be recast as follows.

$$\left\| \boldsymbol{F}_{opt} - \boldsymbol{F}_S\, \boldsymbol{F}_{PS} \boldsymbol{F}_{BB} \right\|_F^2$$

Let $A = \boldsymbol{F}_{opt} - \boldsymbol{F}_S\, \boldsymbol{F}_{PS} \boldsymbol{F}_{BB}$.

$$\emptyset = \|A\|_F^2 = A : A$$

$$\begin{aligned} d\emptyset &= 2A{:}\,\boldsymbol{F}_S d\,\boldsymbol{F}_{PS}\boldsymbol{F}_{BB} \\ &= 2(\boldsymbol{F}_S^{\mathrm{H}} A \boldsymbol{F}_{BB}^{\mathrm{H}}) : d\,\boldsymbol{F}_{PS} \\ \frac{\partial \emptyset}{\partial \boldsymbol{F}_{PS}} &= 2(\boldsymbol{F}_S^{\mathrm{H}} A \boldsymbol{F}_{BB}^{\mathrm{H}}) \\ &= 2\left(\boldsymbol{F}_S^{\mathrm{H}} \boldsymbol{F}_{opt} \boldsymbol{F}_{BB}^{\mathrm{H}} - \boldsymbol{F}_S^{\mathrm{H}} \boldsymbol{F}_S\, \boldsymbol{F}_{PS} \boldsymbol{F}_{BB} \boldsymbol{F}_{BB}^{\mathrm{H}}\right) \end{aligned}$$

By setting the gradient to zero we have

$$\boldsymbol{F}_S^{\mathrm{H}} \boldsymbol{F}_{opt} \boldsymbol{F}_{BB}^{\mathrm{H}} = \boldsymbol{F}_S^{\mathrm{H}} \boldsymbol{F}_S\, \boldsymbol{F}_{PS} \boldsymbol{F}_{BB} \boldsymbol{F}_{BB}^{\mathrm{H}} \quad (22)$$

By adding $\boldsymbol{F}_{PS}$ to both sides of (22), equation (15) will be determined.

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